A FREQUENCY-DIVISION MULTIPLEX SYSTEM FOR USE IN SHIPBOARD INTERNAL VOICE COMMUNICATIONS

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THESIS

A FREQUENCY-DIVISION MULTIPLEX SYSTEM
FOR USE IN
SHIPBOARD INTERNAL VOICE COMMUNICATIONS

by

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(20. ABSTRACT Continued)

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A Frequency-Division Multiplex System For Use In Shipboard Internal Voice Communications

by

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Lieutenant Commander, United States Navy
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requirement for the degree of

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ABSTRACT

A voice communications system that employs frequency-division multiplexing to add four voice channels to existing hardwired, audio-range circuits, is described. Phase-locked loop technology is used in the project, utilizing this new class of monolithic circuit. A study of an existing sound-powered telephone system is made and the compatibility of the multiplexing scheme within this system is investigated.

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I. INTRODUCTION

The increased sophistication of present day weapons systems has raised the demand for additional data transfer capabilities on board many ships. The classic and most expensive technique is to install more wire for each new circuit. The ability to utilize fully the existing frequency spectrum capability of installed wire circuits will greatly reduce the cost of new installations of communications and data handling systems.

A voice communication system utilizing the bandwidth between fifty and two hundred kilohertz was designed, built and tested. This system would supplement the existing voice circuits and add tremendous communications flexibility. The system uses frequency modulation for transmission and multiplexes four channels above the audio range. It was designed with the intent that operation would be on a not-to-interfere basis with existing circuits. The variable impedance and loss with frequency characteristics of these existing circuits makes the use of a frequency modulation-demodulation scheme attractive.[1]

The system was designed with four stations on each of four channels with each station able to signal any other

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[1]Naval Air Engineering Center WTA Project 3527-001, Preliminary Aviation Weapon Movement Control System Definition, Systems Division, Washington Technological Associates, Inc., Section III, 14 February 1972.

including stations on other channels. Cross-channel communications is possible, but it is envisioned that "most frequent customers" would use the same channel.

Phase-locked loop technology was used extensively in this system. The development of a complete, single-chip phase-locked loop has made attractive their application for FM modulation-demodulation schemes. A single packaged device with a few external components offers all the benefits of phase-locked loop operation including independent center frequency and bandwidth adjustment, high noise immunity, high selectivity, high-frequency operation, and center-frequency tuning by means of a single external component. The appendix contains a glossary of terms pertinent to phase-locked loop operation and a brief discourse on the theory of their operation.

One of the major concerns in developing this system is its compatibility with the audio-range circuit using the same pair of wires. The sound-powered circuit produces effects on the prototype model that have to be accommodated. A section of this work will be devoted to a discussion of these problems and their solutions.

The major components of the system, modulator, demodulator, tone generator and tone decoder, all have somewhat different design parameters. Therefore a separate section of this paper will encompass each major component and its design. System design and fabrication will be discussed in other sections. Conclusions and desirable modifications for future systems are in the last section.

II. SYSTEM DESIGN

The intent of this thesis was to design, build and test a prototype modem system to supplement existing hard-wired communications circuits on board naval vessels. Special emphasis was placed on extending the capability of the sound-powered circuits that permeate all naval vessels.

The overriding principle followed in the design of this system was simplicity. It was felt that if this system were to be taken into production that the reliability and cost should be compatible to that of the existing sound-powered communications equipment. In this regard the prototype was designed with the intent that no new headset would be incorporated. Thus the planned system would utilize existing sound-powered circuits and headsets, and would be a "black box" with input and output ports between the telephone and the line. A system block diagram is shown in Figure 1.

In selecting the method of modulation and the band of frequencies to be utilized for this system the major parameter considered was the transmission and impedance versus frequency characteristics of the line. No actual tests were performed by the author, and most information on these characteristics of sound-powered lines was taken from existing literature.[1]

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[1]Naval Air Engineering Center WTA Project 3527-001, Preliminary Aviation Weapon Movement Control System Definition, System Division, Washington Technological Associates, Inc., p. II-2, 14 February 1972.

SYSTEM BLOCK DIAGRAM

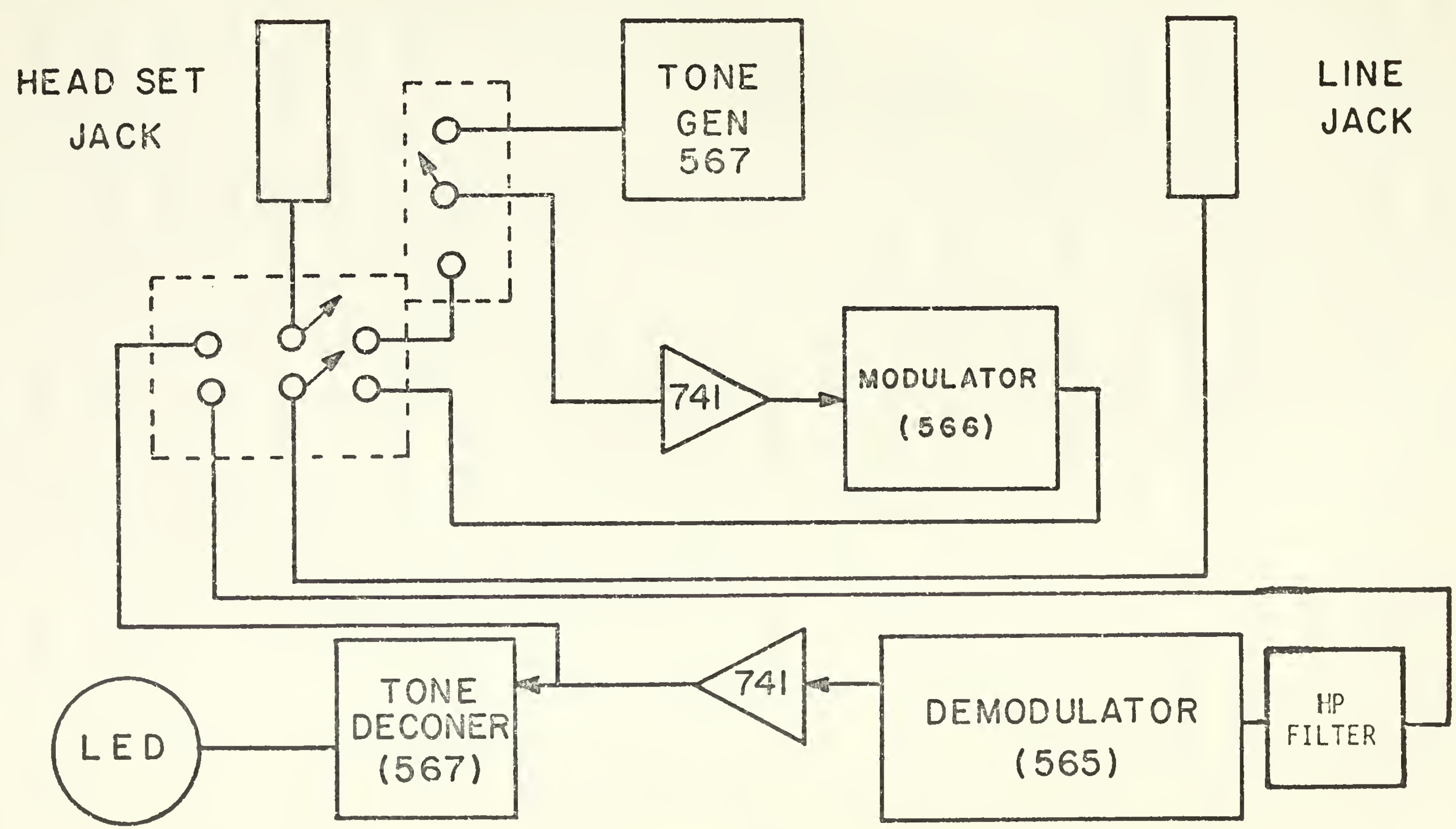

FIGURE I

The variable loading conditions caused by the adding or subtracting of stations on the sound-powered line produce a variable line impedance. Because of this characteristic frequency modulation was chosen as the scheme for information transfer.

Line losses increase rapidly above two hundred kilohertz, therefore the band of frequencies between fifty and two hundred kilohertz was chosen for this system. The capture range of the demodulator determines the bandwidth of each channel. It is sixty percent of the free-running frequency without external adjustment. It can be adjusted to twenty percent of the free-running frequency with the addition of a resistor that limits the output voltage levels. Thus to keep the demodulator on one channel from capturing signals from adjacent channels, and considering the bandwidth available, the number of channels was limited to four.

Channel selection is accomplished by picking a particular frequency determining resistor for both the modulator and demodulator. A double-pole, four-position rotor switch selects these resistors and thus becomes the channel selector switch. The same process is used for selecting the tone generator frequency, using a single-pole, five-position switch which is presented on the face of the box as the station selector switch. The fifth position on the station selector switch presents an open circuit between the tone generator and its +12 volt power supply, shutting it off.

This "rest" position was incorporated to save power and prevent feed through in the signaling switch when it is not in use.

A press-to-talk switch was incorporated so that the carrier frequency oscillator is off in the "listen" position. This prevents distortion of the incoming signal at the line jack.

A. MODULATOR

The basic component used in the construction of the modulator was a type 566 integrated-circuit function generator. The 566 is a voltage-controlled oscillator of high stability and linearity. The frequency of oscillation is determined by an external resistor and capacitor and by the bias voltage applied to the control terminal. In this application a bias voltage of 10.5 volts was chosen which put the operating point midway on the frequency versus control voltage curve specified in the design literature.[1] This also meets one of the specific design criteria of the 566 in that the control terminal must be biased with a voltage (V_c) in the range

$$\frac{3}{4} V^+ \leq V_c \leq V^+$$

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[1] Signetics Corporation, Linear Integrated Circuits, Vol. 1, p. 216, 1972.

where V^+ is the total supply voltage of 12 volts. In Figure 2 this control voltage is set by the voltage divider formed with 15K and 100K resistors. The external resistor controlling the frequency of oscillation, R_1 in Figure 2, was replaced with a bank of four variable resistors, each with a value that sets the modulator frequency for a specific channel. Each of these resistors was wired through a rotor switch which thus becomes the channel selector switch. This center frequency (f_o) is given approximately by

$$f_o \cong \frac{2(V^+ - V_c)}{R_1 C_1 V^+}$$

The value of C_1 was chosen such that the four channel frequencies could be selected by changing R_1 and still meet the design criteria of the 566 that R_1 be in the range

$$2K\Omega \leq R_1 \leq 20K\Omega$$

This was accomplished by choosing a 0.001 microfarad capacitor for C_1.

A capacitor of 0.001 microfarad value was placed between pins 5 and 6 to eliminate oscillations in the control current source within the device.

The available outputs are taken at pins 3 and 4. The triangular output from pin 4 is passed through a simple RC filter to eliminate harmonics. This signal is routed through a press-to-talk switch to the line output jack.

MODULATOR

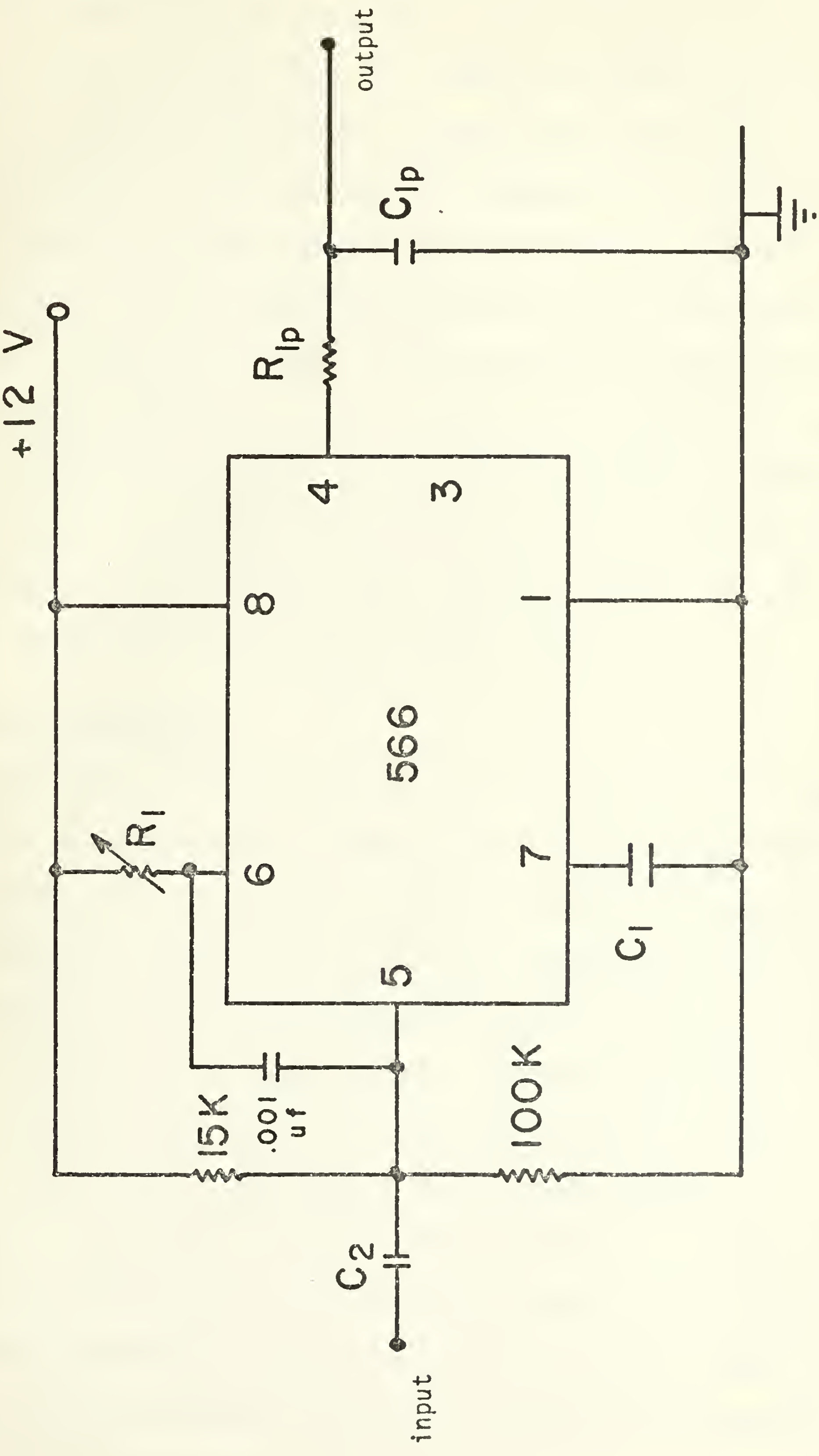

FIGURE 2

The input from the headset jack is routed through the press-to-talk switch and the signal generator switch to ensure that the microphone is dead when signaling. This prevents any loading by the microphone on the signal generator. Thus for signaling to occur the signaling switch is thrown while the press-to-talk switch is in the "talk" position.

The output amplitude of the modulator drops off with increasing frequency. In the range of frequencies used this variation was from 0.7 volts on the lowest channel to 0.2 volts on the highest. No circuit was found to be necessary for automatic level control due to the wide amplitude tolerance of the phase-locked loops.

B. TONE GENERATOR

The basic component of the signal generator is again the 566 with the same biasing and external design parameters. R_1 again was replaced by a bank of variable resistors and C_1 chosen to include all signaling frequencies and still maintain R_1 within the previously stated design parameters for the 566 device. This system incorporates four stations per channel, thus only four separate tones are required to signal all other stations of the sixteen stations of the system. The output of the tone generator is routed through the station selector switch to the input of the modulator. Thus the signaling tone modulates the carrier signal and there are no frequency components within the frequency range used by the sound-powered telephones.

The tone generator frequencies were chosen to be 4000, 5000, 6000 and 7000 hertz. These frequencies are audible on the headset line after demodulation, but allow the signal decoder to be set above normal voice range activation. This eliminates spurious decoding of voice signals that could be misconstrued as signaling tones.

C. DEMODULATOR

The basic component of the demodulator is the 565 phase-locked loop integrated circuit. The 565 is a general-purpose circuit designed for highly linear FM demodulation. During lock, the average dc level of the phase comparator output signal is directly proportional to the frequency of the input signal. As the input frequency shifts, it is this output signal which causes the VCO to shift its frequency to match that of the input. Consequently, the linearity of the phase comparator output with frequency is determined by the voltage-to-frequency transfer function of the VCO.

The connection diagram is shown in Figure 3. The VCO free-running frequency is given approximately by

$$f \cong \frac{1}{4\ R_1\ C_1}$$

R_1 was replaced by a bank of four variable resistors each setting the free-running frequency of the VCO to that required for its specific channel. Normally, this free-running frequency is centered within the range of expected input

DEMODULATOR

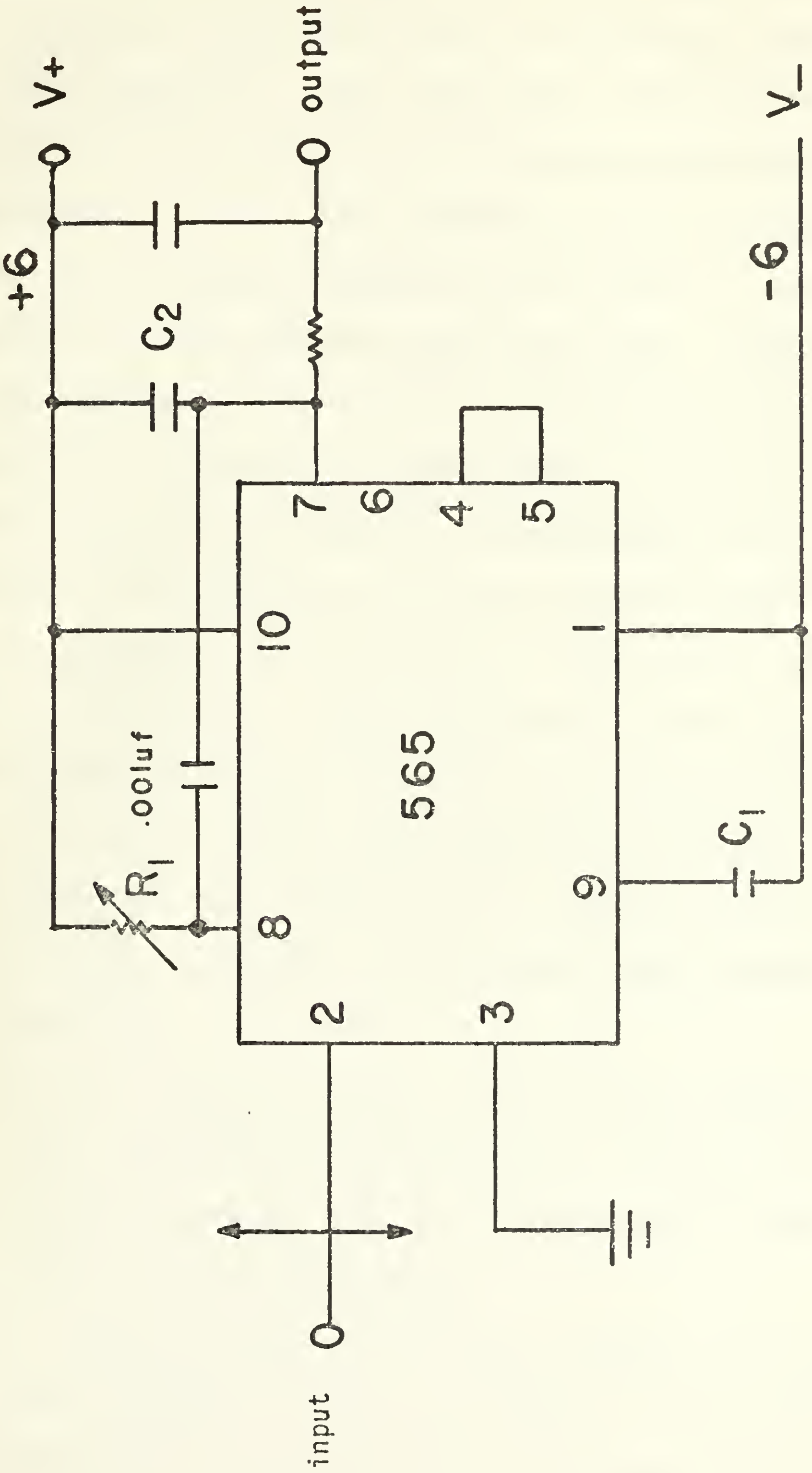

FIGURE 3

frequencies. In this application, the free-running frequency for channels 1 and 4 was offset to provide a guard band from the other channels. For channel one the carrier frequency is fifty kilohertz but the free-running frequency was set at 45 kilohertz. Thus, for a capture range of twenty percent of the free-running frequency, or 45 kHz ± 9 kHz the highest frequency in the capture range is 54 kHz. Channel two free-running frequency was centered on 75 kHz giving a capture range of 60-90 kHz, thus providing a 6 kHz guard band between channels one and two. The channel three carrier frequency was adjusted to 120 kHz and the demodulator free-running frequency was centered on this value, giving a capture range of 96-144 kHz. This provides a 6 kHz guard band between channels two and three. Channel four carrier frequency was set at 155 kHz, but the demodulator free-running frequency was centered at 187.5 kHz giving a capture range of 150-225 kHz and again providing a 6 kHz guard band between channels three and four.

The off centering of the free-running frequency of the VCO in the demodulator increases the pull-in time for lock, but the off centering used meets the capture range requirements of the design literature.

The incoming signal was routed through the press-to-talk switch and coupled through a capacitor to the 565 demodulator. The loop filter is a double-pole RC network using the 3.6K internal resistor, two external 0.15 microfarad capacitors, and another 3.6K external resistor.

Pins 2 and 3 were grounded for dc through 4.3K resistors, assuring no dc voltage difference between the pins, a design requirement.[1] Pins 4 and 5 are shorted to connect the VCO and the phase comparator within the device. A 0.001 microfarad capacitor is connected between pins 7 and 8 to eliminate any possible oscillations in the control current source. The output of the demodulator taken at pin 7 is coupled through a capacitor to a 741 operational-amplifier integrated circuit with a voltage gain of ten. This output was routed to the headset jack and to the input of the tone decoder.

D. TONE DECODER

The basic component of the tone decoder is the 567 phase-locked loop integrated circuit. The 567 is designed specifically for frequency sensing or tone decoding. It has a controlled oscillator, phase detector and a second auxiliary or quadrature phase detector like otherphase-locked loops. In addition, however, it contains a power output stage which is driven directly by the quadrature phase detector output. During lock, the quadrature phase detector drives the output stage on, so the device functions as a tone decoder. The tone decoder center frequency and bandwidth are determined by the center frequency and capture

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[1]Signetics Corporation, <u>Linear Integrated Circuits</u>, Vol. I, p. 212, 1972.

range of the loop portion of the device. Since a tone decoder, by definition, responds to a stable frequency, the lock or tracking range is relatively unimportant. In other words, if the decoder cannot capture the tone signal it of course cannot track it, but the spread of the tracking range determines the capture range of the device.

The connection diagram for the tone decoder is shown in Figure 4. The values of R_1 and C_1 which set the center frequency were chosen for the four different center frequencies, 4000, 5000, 6000 and 7000 Hz, corresponding to four different stations. These frequencies were selected to produce audible tones at the headset jack during signaling and the tone decoder is protected from normal speech signal activation occurring at 3000 Hz and below. The capacitor C_2 was selected to provide for a five percent bandwidth about the center frequency. C_3, which sets the band edge of a low-pass filter which attenuates frequencies outside the detection band, was made equal to twice C_2. This is a compromise between attenuation of spurious output and turn-on time of the output stage transistor.

The primary output of the tone decoder is taken at pin eight which is the collector of a transistor within the device and is normally at +6 volts potential. When an in-band signal is present this transistor saturates; its collector voltage being less than one volt at full output current. This output drives a load consisting of an LED

TONE DECODER

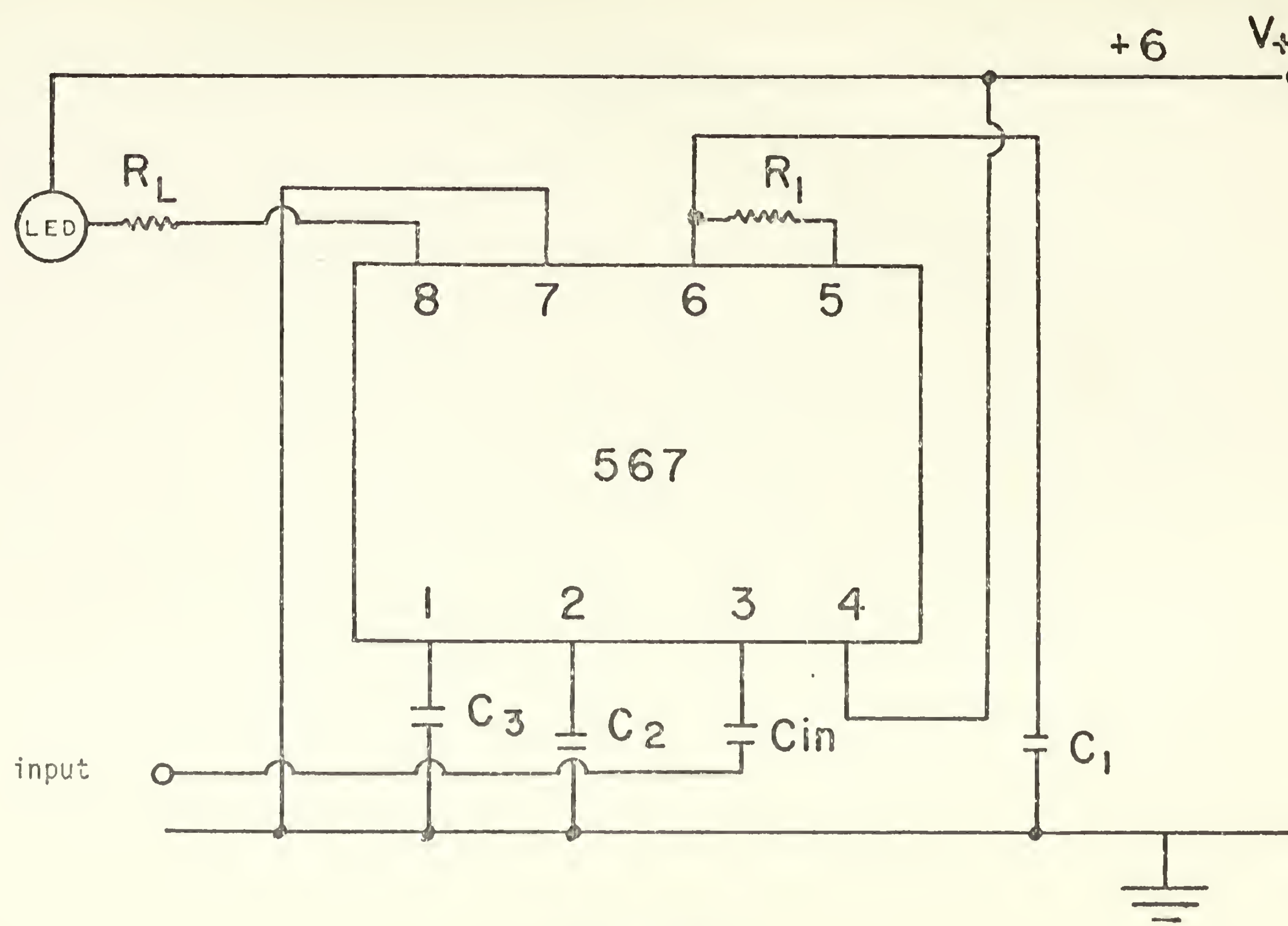


FIGURE 4

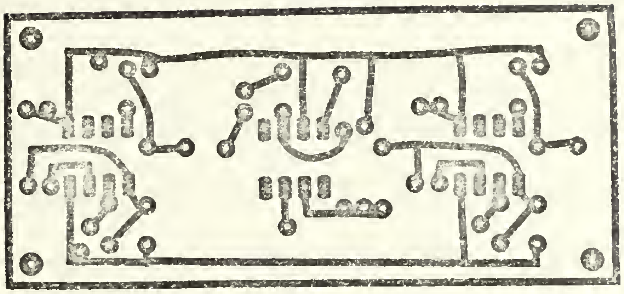
in series with a 100 ohm current-limiting resistor. Thus an in-band signal tone will be audible in the earpiece and will activate the LED on top of the box.

III. FABRICATION

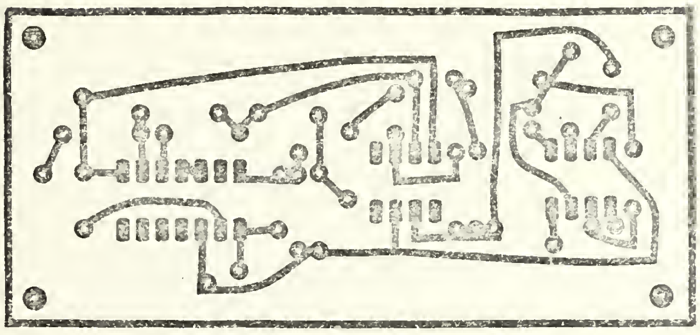
The tone generator, microphone amplifier, and the demodulator are placed on the integrated circuit board shown in the top view of Figure 5. No serious attempt was made at minaturization because it was felt more important to have the various test points readily accessible during development. The integrated circuits used are of the eight-pin duel - in line - package variety which provide a low profile and easy placement on the integrated circuit board.

The demodulator, final amplifier, and the tone decoder are placed on the integrated circuit board shown in the bottom view of Figure 5. These two sections are mounted to the top of a $3\frac{1}{2}$ by 6 inch box with $\frac{1}{2}$ inch spacing from the top to allow for wiring. The three banks of ten kilohm variable resistors are individually mounted on their own integrated circuit board and fastened on top of the other two boards. Thus the entire device is mounted to the top of the box making entry for testing or maintenance easy. The wire bundles from the resistor banks servicing the modulator and demodulator are routed to the two-pole, four position rotor switch. The bank servicing the tone generator is routed to a single-pole, five-position rotor switch.

The telephone and line jacks used are standard miniature microphone jacks, from which the signals are routed through the two-pole, double-throw press-to-talk switch to the various input and output points of the devices.



(a) Top View



(b) Bottom View

FIGURE 5. INTEGRATED CIRCUIT BOARD

Outside power is provided through four standard banana jacks to the four major buses of the system: +12 volts, +6 volts, -6 volts and the ground bus. Because the carrier frequency is a fairly precise function of the applied bias voltage it was considered mandatory to use regulated laboratory power in the development of the system. Battery power, certainly a desirable alternative, would require a regulator circuit, and was left for future development.

The component count includes the three major phase-locked loop integrated circuits, the 565, 566 (2), and 567, and the 741 (2) operational amplifier, plus the various resistors and capacitors. The total list is given in the appendix.

IV. COMPATIBILITY

A. SOUND-POWERED HEADSET

In the study of the impedance characteristics of the sound-powered telephone line it was found that as more headsets were added to the line, the impedance of the line steadily decreased with increasing frequency.[1] In examining this response for an individual sound-powered headset it was found that the impedance decreases to about sixteen ohms and remains at that value from about 8 kHz to 200 kHz. The response was measured by inputting a 1 volt signal from a Wavetek 142 signal generator into the sound-powered headset in series with a 1000 ohm resistor, and measuring the voltage across this resistor with a Hewlett-Packard 427A voltmeter. This voltage is shown in Figure 6.

This indicates that at the frequencies used in the prototype a low impedance will be seen by the modulator carrier signal whenever sound-powered phones are connected to the line. In laboratory tests it was found that this low load impedance resulted in amplitude distortion with resulting harmonics.

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[1]Naval Air Engineering Center WTA project 3527-001 Preliminary Aviation Weapon Movement Control System Definition, Systems Division, Washington Technological Associates, Inc., Table II-2, 14 Feb. 1972.

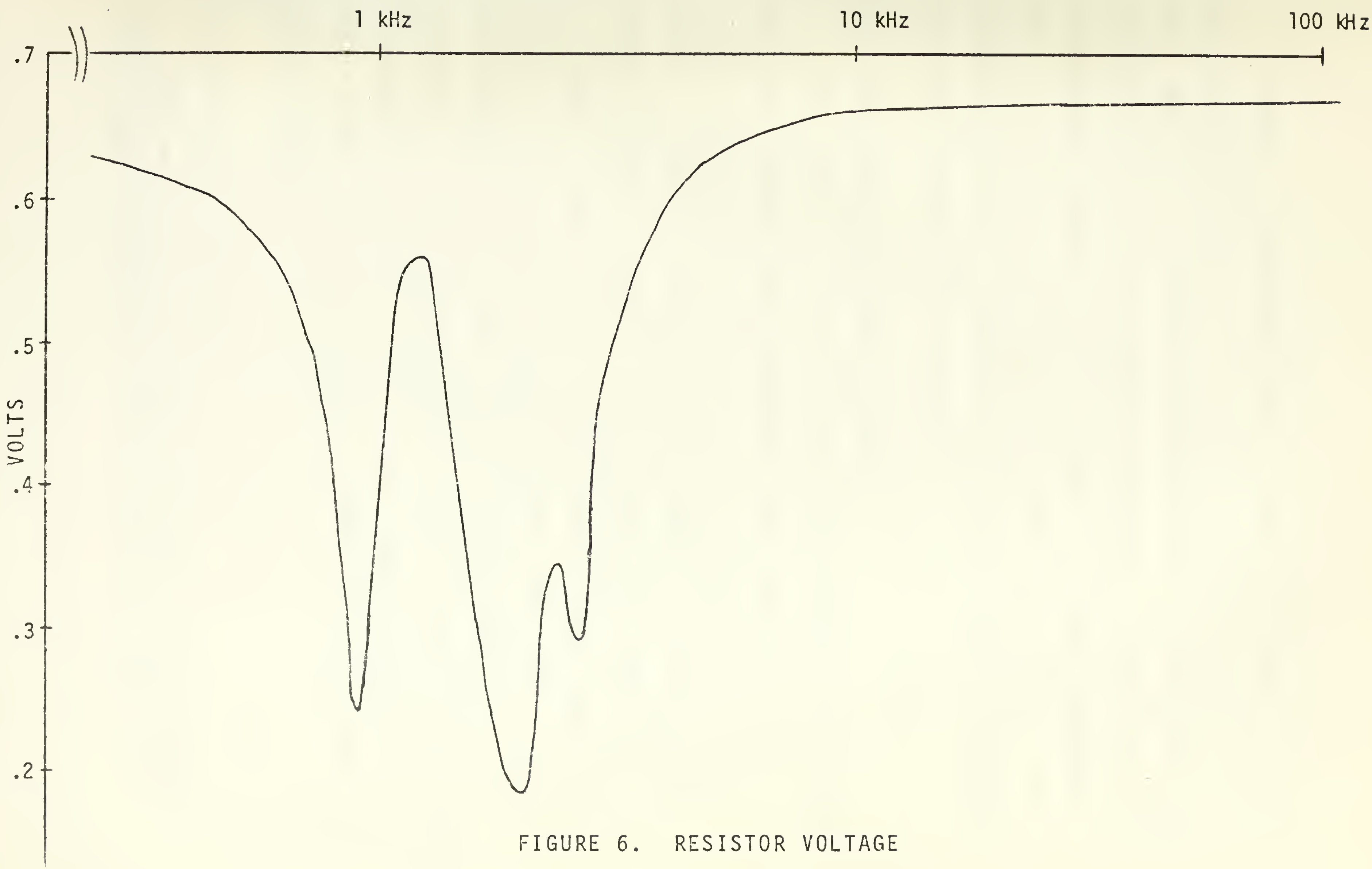

FIGURE 6. RESISTOR VOLTAGE

To test for intermodulation distortion produced by the sound-powered phones, 60 and 120 kHz carrier signals of 1 volt amplitude frequency modulated with a 3 kHz signal were placed directly on the headset terminals. On the two types of headsets tested the 3 kHz signal was clearly audible.[1] This was true for both 60 and 120 kHz input signals. The headset acts like a demodulator, and both types of headsets have this inherent non-linearity.

B. CONCLUSIONS AND MODIFICATIONS

When tested in the laboratory the prototype users could hear the sound-powered telephone users and vice versa. Indeed not a good situation for a proposed multiplexing system. The two occurances are caused by two different events and two different modifications were required.

When an audio signal was combined with the carrier signal at the input of the demodulator, the audio signal fed through to the output and was clearly audible. The solution to this problem was to insert an active second order Butterworth high-pass filter at the input of the demodulator. This filter, shown in Figure 7, effectively eliminated the feed through.

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[1]The two types of handsets used were: Sound-Powered Handset, manufactured by Wheeler Electronic Corp. and a RCA Handset, MI-2040-A.

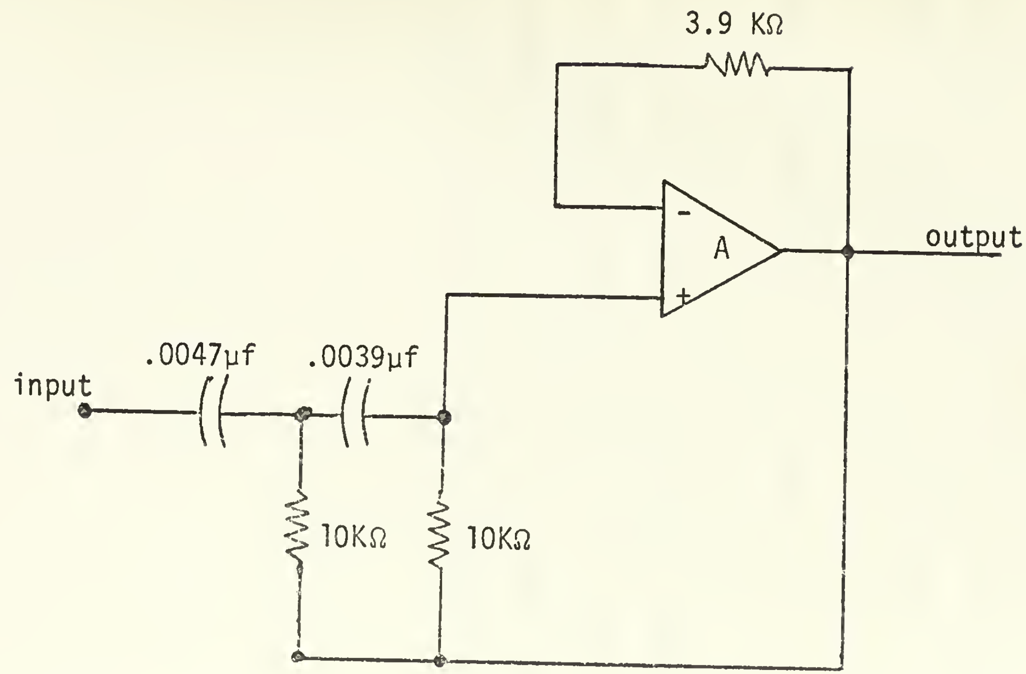

FIGURE 7. HIGH-PASS FILTER

The second problem, the audio-range users hearing the prototype, presents a more difficult situation. The problem is caused by non linearity in the sound-powered telephone which causes detection of the frequency-modulated carrier. Two solutions are obvious. One would be to put a low-pass filter at the input of each sound-powered telephone. A filter was tried which gave good results. The difficulty is that it would be very undesirable to have to introduce such a circuit at each sound-powered outlet in the system. A second possibility is to reduce the level of the carrier signals on the line by 20 to 30 dB. This would require an increase in the gain of the input amplifier to compensate for the reduced signal level. Time did not permit implementation of this solution, but there should be no great difficulty involved, and this is the recommended solution.

V. FUTURE DEVELOPMENT

This project is only one attempt at the solution of better utilizing the frequency spectrum capability of existing wire circuits. As the system was taken through the various stages of development different design ideas were raised, and during testing some deficiencies noted. The significant ones are detailed in this section.

A. SYSTEM CHANGES

In the prototype system the electronics were housed in a $3\frac{1}{2}$ x 6 inch box. A better form of packaging would house the entire unit within the headset. This would require miniaturization of the package, especially the switching, which takes up to fifty percent of the physical space in the prototype model. A portable unit would also require battery power, which because the carrier frequency of the modulator is a precise function of applied bias voltage, would mean the addition of a regulation circuit. Waste space would have to be eliminated from the integrated circuit boards. The single-pole press-to-talk switch of the headset would have to be integrated with the switch on the black box to do away with the two switches used in the prototype system.

In designing this headset-packaged model consideration should be given to incorporating better performing microphones and earphones. There are many miniature models available

that would reduce the space taken up by the present standard sound-powered units.

Additional channels and stations might be desired for certain applications. This requires a smaller capture range for each channel in the demodulator and for each station in the tone decoder. Decreasing the capture range attenuates the output amplitude, thus additional amplification might be required. Increasing the number of channels within the design bandwidth does not cause any increase in the amplitude variation of the output. Thus an automatic leveling circuit is still not indicated.

During signaling with the prototype model all four stations on one channel receive the audible tone. The lighted LED solves the ambiguity of which station is being called, but the light is not attention getting. The addition of a buzzer device is indicated which would work off of the tone decoder as does the LED. Miniature, low powered buzzer devices are available, and their incorporation would not add significantly to power consumption or take up excessive physical space.

The prototype boxes were set up with the station, determined by the center frequency of the tone decoder, hard-wired within the box. The addition of a bank of resistors, wired through a selector switch would add the capability of changing the designation of any station. This would add considerable flexibility over the prototype model.

Plugging the prototype system into a sound-powered outlet precludes the use of that outlet in the normal manner. This inconvenience could be circumvented by incorporating a jack for the sound-powered system on the black box. Alternately the carrier outlets could be connected through the terminals of the sound-powered system rather than being plugged into existing outlets.

B. COMPONENT CHANGES

The present prototype system layout requires too much hand wiring. The integrated circuit board could be better designed to do away with long power supply and ground lines, resistor bank wire bundles and the many leads to the various switches.

The power supply lines as incorporated require four different inputs, +6 and -6 volts for some boards, and +12 for others. A redesign could certainly reduce the number of voltages required from three to two.

APPENDIX A

A. PHASE-LOCKED LOOP OPERATION

The phase-locked loop is a feedback system comprised of a phase comparator, a low-pass filter and an error amplifier in the forward signal path, and a voltage-controlled oscillator (VCO) in the feedback path. The block diagram of a basic PLL system is shown in Figure 8. Perhaps the single most important point to realize when designing with the PLL is that it is a feedback system, and hence is characterized mathematically by the same equations that apply to other, more conventional feedback systems. The parameters in the equations are somewhat different, however, since the feedback error signal in the phase-locked system is a phase, rather than a current or voltage signal as is usually the case in conventional feedback systems.

A rigorous mathematical analysis of the system is quite cumbersome and will not be repeated here.[1] However, from a qualitative point of view, the basic principle of PLL operation can be briefly explained as follows: with no input signal applied to the system, the error voltage, V_d, is

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[1] Signetics Corporation, <u>Digital-Linear-Mos Applications</u>, pages 6-6 - 6-11, 1973.

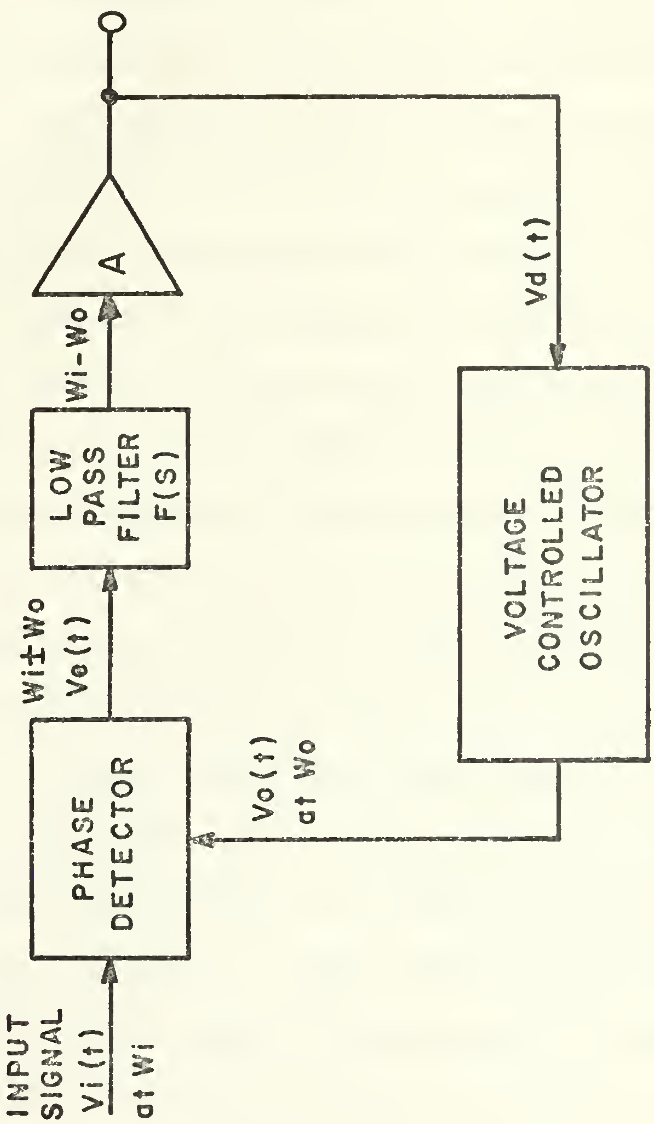
INPUT SIGNAL Vi(t) at Wi
PHASE DETECTOR
Wi±Wo Ve(t)
LOW PASS FILTER F(S)
Wi-Wo
A
Vd(t)
VOLTAGE CONTROLLED OSCILLATOR
Vo(t) at Wo

FIGURE 8

equal to zero. The VCO operates at a set frequency, ω_o, which is known as the free-running frequency. If an input signal is applied to the system, the phase comparator compares the phase and the frequency of the input with the VCO frequency and generates an error voltage, $V_e(t)$, that is related to the phase and the frequency difference between the two signals. This error voltage is then filtered, amplified, and applied to the control terminal of the VCO in such a manner as to force the VCO frequency to change in a direction that reduces the frequency difference between ω_o and the input signal. If the input frequency, ω_i, is sufficiently close to ω_o, the feedback causes the VCO to synchronize or lock with the incoming signal. Once in lock, the VCO signal is identical to the input signal except for a finite phase difference. This net phase difference, θ_o, is necessary to generate the corrective error voltage, V_d, to shift the VCO frequency from its free-running value to the input signal frequency, ω_i, and thus keep the PLL in lock. The self-correcting ability of the system also allows the PLL, once it is locked, to track the frequency changes of the input signal. The range of frequencies over which the PLL can maintain lock with an input signal is defined as the "lock range" of the system. The band of frequencies over which the PLL can acquire lock with an incoming signal is known as the "capture range" of the system and is never greater than the lock range.

Another means of describing the operation of the PLL is to observe that the phase comparator is in actuality a multiplier circuit that mixes the input signal with the VCO signal. This process produces the sum and difference frequencies $\omega_i \pm \omega_o$. When the loop is in lock, the VCO duplicates the input frequency so that the difference frequency component, $(\omega_i - \omega_o)$, is zero. In this case the output of the phase comparator contains a dc component. This dc component is then amplified and fed back to the VCO. Notice that when the loop is in lock, the difference frequency component is always a dc voltage, so that lock range is independent of the band edge of the low-pass filter.

Consider now the case where the loop is not yet in lock. The phase comparator again mixes the input and VCO signals to produce sum and difference frequency components. Now, however, the difference component may fall outside the band edge of the low-pass filter and be removed along with the sum frequency component. If this is the case, no information is transmitted around the loop and the VCO remains at its initial free-running frequency. As the input frequency approaches that of the VCO, the frequency of the difference component decreases and approaches the band edge of the low-pass filter. Now some of the difference component is passed, which tends to drive the VCO towards the frequency of the input signal. This, in turn, decreases the frequency of the difference component and allows more information to

be transmitted through the low-pass filter to the VCO. This is essentially a positive feedback mechanism which causes the VCO to snap into lock with the input signal. With this mechanism in mind, the term "capture range" can again be defined as the frequency range centered about the VCO initial free-running frequency over which the loop can acquire lock with the input signal. The capture range is a measure of how close the input signal must be in frequency to that of the VCO to acquire lock. The capture range can assume any value within the lock range and depends primarily upon the band edge of the low-pass filter together with the closed-loop gain of the system. It is this signal capturing phenomenon which gives the loop its frequency selective properties.

It is important to distinguish the capture range from the lock range. The tracking range can be defined as the frequency range usually centered about the VCO free-running frequency over which the loop can track the input signal once lock has been achieved.

When the loop is in lock, the difference-frequency component on the output of the phase comparator (error voltage) is a dc voltage, and will always be passed by the low-pass filter. Thus, the tracking range is limited by the range of error voltage that can be generated and the resulting VCO frequency deviation. The tracking range is essentially a dc parameter and is not affected by the band edge of the low-pass filter.

B. DEFINITION OF TERMS

FREE-RUNNING FREQUENCY

Frequency of VCO without input signal, both inputs grounded.

CAPTURE RANGE

That range of frequencies in the vicinity of the free-running frequency over which the loop will acquire lock with an input signal initially starting out of lock.

LOCK RANGE OR TRACKING RANGE

That range of frequencies in the vicinity of the free-running frequency over which the VCO, once locked to the input signal, will remain locked.

CENTER FREQUENCY

The free-running frequency of the voltage controlled oscillator (VCO) in the absence of an input signal.

COMPONENTS

NAME	QUANTITY
741 op amp	2
565 PLL	1
566 VCO	2
567 PLL	1
LED and holder	1
Mini jack	2
Rotor switch (2p/4pos)	1
Rotor switch (1p/5pos)	1
2P2T toggle switch	2
"Banana jack"	4
10K variable resistor	12
container	1
CAPACITORS:	
.001 microfarad	7
.01 "	2
.02 "	2
.0068 "	1
.15 "	2
.47 "	1
1 "	1
10 "	1
33 "	1
RESISTORS:	
10K	3
100K	4
4.7K	2

One each of the following: 16K, 1.5K, 51K, 200K, 1.2K, 15K, 3.9K and 100 ohm.

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